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(71) Applicant *(for all designated States except US)*: DUNE S.R.L. [IT/IT]; Via Tracia, 4, I-00183 Rome (IT).

(72) Inventor; and
(75) Inventor/Applicant *(for US only)*: GASPARINI, Otello [IT/IT]; Piazza Vespri Sicilini, 17, I-00162 Rome (IT).

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(54) Title: SPECTRAL PROCESSOR FOR THIRD GENERATION CELLULAR TELEPHONY

(57) Abstract: Third generation (3G) mobile telephony systems, such as UMTS (Universal Mobile Telecommunications Services), use DS-CDMA (Direct Sequence-Coded Division Multiple Access). The *base station* transmits simultaneously the aggregated sum of the symbol streams to all mobile users in the cell. The propagation channel, affected by multiple propagation paths and associated delays, generates at each mobile (ME) auto-interference from delayed replicas of its symbol stream and from streams to other mobiles in the cell. The invention is a set of new signal processors, indicated collectively as CPE processor, used in reception at the mobile for symbol estimation, to improve downlink communication by minimizing Inter Symbol Interference (ISI) and intracell Multi User Interference (MUI). Basically the CPE processor is a channel equaliser, followed by code correlation and detection of the symbols stream addressed to the ME.

1 DESCRIPTION

TITLE: SPECTRAL PROCESSOR FOR THIRD GENERATION CELLULAR TELEPHONY

1.1 Technical Field

Mobile Wireless Communication

1.2 Background art

DS-CDMA codes of the mobile users in the cell are mutually orthogonal (eq. (1)), that is when, in the absence of multipath an user *correlates* the received signal with its own code, the symbol streams addressed to other users, that potentially may interfere with the user, are cancelled.

The antenna in the radio-base can employ one or more elements, while today the mobile user employs mostly one element, in SISO (Single Input Single Output) or MISO in downlink (Multiple Input Single Output) configurations.

In a cell the orthogonality between user codes is maintained in reception only for an ideal propagation channel [1].

For non ideal (dispersive) channels, affected by multiple propagation paths and associated delays, the orthogonality is lost due to the time dispersion of the transmitted signal.

A dispersive channel causes auto-interference between delayed replicas of a user symbol stream (ISI, Inter Symbol Interference) and interference from streams relative to other users in the cell (MUI, Multiple User Interference). In the first generation UMTS, it is foreseen to employ mostly a simple processor (Rake type [1]). This is basically a matched filter collecting delayed replicas of the symbol stream transmitted to the intended user; minimising the interference from the streams addressed to other users is not provided.

Other types of processors, indicated as MUD (Multi User Detector) [1] [2], would be able to cope with both auto-interference and with interference from other users. Todays MUDs are complex to implement, they are expensive and require a considerable energy consumption and battery autonomy of the mobile receiver.

With respect to simple processors such as Rake, MUD processors would permit alternatively

- to increase, with the same power transmitted from the radio-base, the maximum communication range between radio-base and user, that is the cell dimension;
- to maintain the same performances with less power in the transmission

Actually, the trend in decreasing MUD complexity is slow, and in the medium term it is likely that sub optimal processors are used, whose performances are lower than MUD.

1.3 Disclosure of the invention

The invention consists of a set of signal processors, indicated collectively as CPE processor, used in reception at the mobile (ME) for symbol stream estimation; the scope is to minimize Inter Symbol Interference (ISI) and intracell Multi User Interference (MUI). Basically the processor is a novel equaliser of the channel linking the base station and a mobile user (ME), followed by code correlation and detection of the symbols stream addressed to the ME.

The processor is significantly less complex than nowadays MUD, and lower implementation cost for the mobile receiver.

The first CPE [Cyclic Prefix Extension] processor we will consider is based on the introduction of a time extension of the signal transmitted from the radio-base; the extension is implemented with a Cyclic Prefix [3] (see fig.1), which is a kind of guard time interval between subsequent symbols.

A DS-CDMA code associated to a symbol to be transmitted is a sequence of n time elements (chips), which can be represented by a column vector $\mathbf{C}$ with n rows and one column; a vector element is the same as a time code element. The *Fast Fourier Transform* (FFT) of this vector, implemented at chip rate with respect to discrete time samples, is also a vector $\mathbf{T}$, having the same number of elements as $\mathbf{C}$.

The signal at the radio base transmitter, which is the aggregated sum of all the codes, each multiplied by the symbol addressed to a different user in the cell, comprises n chips before the extension; for implementing a Cyclic Extension, the last d samples are

repeated at the beginning of the n-chips sequence to form the final sequence of $n + d$ samples to be eventually transmitted synchronously from the radio-base.

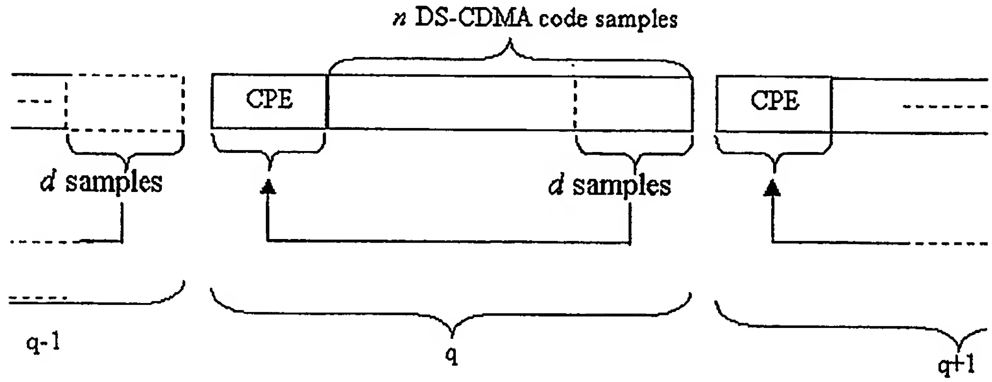

Figure 1: Transmission time extension through Cyclic Prefix (CPE). Subsequent intersymbol intervals are indicated by q-1; q; q+1.

In theory the CPE introduction reduces the spectral efficiency by the factor: (duration of the DS-CDMA code) / (duration of (DS-CDMA + CPE)). The reduction is more than compensated by the achievable advantages. CPE processing is applicable both to SISO and to MISO. A particular *Space Time Block Coding* (STBC) of the transmitted symbol *sequence* concatenated to the DS-CDMA coding, is standardised in UMTS for MISO configuration using two array elements at the radio base (indicated as configuration 2,1) ([1], pag. 97, fig. 6.15; [4]).

The spatial diversity made available by STBC increases the range of communication and its quality and stabilises the performance with respect to channel fluctuation.

1.3.1 Processor for SISO configuration: CPE-1

To simplify the explanation of CPE processors, in downlink (towards users) assume that only two orthogonal DS-CDMA codes are transmitted, associated to two symbols relative to users 1 and 2. The processing can be obviously extended to the transmission of a higher number of codes.

We introduce the notation:

- $(.)^{+}$ conjugation and transposition of a complex vector or matrix ;
- $(.)^{*}$ conjugation of a complex vector or matrix ;
- $(.)^{T}$ transposition of a vector or matrix;

- |.| absolute value of a scalar;
- $\mathbf{C}_1 = (c_{1,1}, c_{1,2}, \ldots, c_{1,n})^T$ is a vector whose components are values of the time chips of the DS-CDMA code for the user 1. The index increases with increasing time. In a similar way vector $\mathbf{C}_2$ for the user 2 is defined; n is the length of the DS-CDMA code;
- s_1, s_2 are the symbols (complex scalars) for the users 1 and 2 associated to codes $\mathbf{C}_1$ and $\mathbf{C}_2$.
- $\mathbf{C}_1\ s_1 + \mathbf{C}_2\ s_2$: transmitted vector, of n chips duration, *before* adding the cyclic prefix;

Let the number of samples (taken at a chip rate) for the coded symbol, *comprising the cyclic extension,* be $n+d$; couples of symbols, for the users 1 and 2, are transmitted every $(n+d)$ samples. In reception, by discarding the d samples corresponding to CPE from the signal received in an interval of $d+n$ samples, the n sample received vector $Y=(y_1, y_2, \ldots, y_n)^T$ is obtained;

The time samples of the channel impulse response, taken at chip rate, are indicated by $h_0, h_1, h_2, \ldots, h_L$ for increasing delay indexes. A channel is non dispersive when only one value is different from zero.

The codes of two users are *orthogonal,* that is their inner product is zero as indicated below:

$$\mathbf{C}_1^{+}\mathbf{C}_2 = \sum_{i=1}^{n} c_{1i}^{*} c_{2i} = 0 \qquad (1)$$

To simplify, receiver noise is neglected in this description

We can write:

$$\mathbf{Y} = \mathbf{H}(\mathbf{C}_1 s_1 + \mathbf{C}_2 s_2) \qquad (2)$$

where **H** is the (n,n) matrix of the channel linking transmitter and *user 1*. Eq. (2) indicates time convolution of the aggregated transmitted signal (at second member in parenthesis) relative to two users, with the impulse channel response (**H**). The convolution is implemented by the propagation channel mechanism, to obtain the vector **Y** (once the cyclic prefix is discarded) received by the user 1. Due to the introduction of the cyclic prefix the channel matrix results *circulant*, that is rows (columns) are cyclic permutations of the first row (column). For example, the channel matrix in the case $n = 5$, with 3 channel samples is

$$\mathbf{H} = \begin{pmatrix} h_0 & 0 & 0 & h_2 & h_1 \\ h_1 & h_0 & 0 & 0 & h_2 \\ h_2 & h_1 & h_0 & 0 & 0 \\ 0 & h_2 & h_1 & h_0 & 0 \\ 0 & 0 & h_2 & h_1 & h_0 \end{pmatrix} \quad (3)$$

A circulant matrix has the following spectral factorisation, **S** and **F** being the diagonal eigenvalue matrix and the unitary eigenvector matrix [5]:

$$\mathbf{H} = \mathbf{FSF}^{+} \quad (4)$$

It is found that matrix **F** does not depend on the channel and performs an *Inverse Fast Fourier Transform* (IFFT: from frequency to time). The F_{rs} element of the matrix **F** is given by [6]

$$\mathrm{F}_{rs} = \frac{1}{\sqrt{n}} e^{j\frac{rs}{n}} \quad (r,s:0,1,\ldots,n-1) \quad (5)$$

S is a diagonal matrix, which is the channel impulse response FFT transform (spectral response), with complex elements. Matrix $\mathbf{F}^{+}$ performs a time to frequency transform. We indicate by $\mathbf{T}_1$, $\mathbf{T}_2$ and **Z** the FFT transforms of the code vectors $\mathbf{C}_1$, $\mathbf{C}_2$ of the two users, and of the received **Y** vector. As $\mathbf{C}_1$, $\mathbf{C}_2$ are orthogonal (eq. (1)), and as matrix **F** is *unitary* (that is $\mathbf{FF}^{+}=\mathbf{F}^{+}\mathbf{F}=\mathbf{I}_n$, where is the identity matrix of dimensions (n,n)) also the transformed vectors $\mathbf{T}_i$ result orthogonal:

$$\mathbf{T}^{+}_1\mathbf{T}_2 = \mathbf{C}^{+}_1\mathbf{FF}^{+}\mathbf{C}_2 = \mathbf{C}^{+}_1\mathbf{C}_2 = 0 \quad (6)$$

Vectors $\mathbf{T}_1$, $\mathbf{T}_2$, indicated as *spectral codes*, do not depend on the received signal and can be calculated beforehand. We can write:

$$\mathbf{Y} = \mathbf{FS}(\mathbf{F}^{+}\mathbf{C}_1 s_1 + \mathbf{F}^{+}\mathbf{C}_2 s_2) \quad (7)$$

By pre-multiplying the previous equation by $\mathbf{F}^{+}$, we obtain:

$$\mathbf{Z} = \mathbf{S}(\mathbf{T}_1 s_1 + \mathbf{T}_2 s_2) \quad (8)$$

Thus, after FFT, the received spectral vector **Z** is the aggregated sum of the spectral vectors of the various users, multiplied frequency-wise by the channel spectral response. Such response is estimated beforehand at the user mobile by processing the receptions relative to certain transmissions with structure known a priori (pilot symbols).

By pre-multiplying both members of (8) by the inverse of the diagonal matrix **S**, we have:

$$\mathbf{S}^{-1}\mathbf{Z} = \mathbf{T}_1 s_1 + \mathbf{T}_2 s_2 \quad (9)$$

The processor of user 1 pre-multiplies the previous equation per T^{+}_{1} to obtain the estimate of the symbol s_1 directed to it; by taking into account (6) we obtain:

(10) $$T_1^{+}S^{-1}Z = s_1$$

We can also write as an alternative to the processing indicated by the eq. (10)

(11) $$C_1^{+}\left(FS^{-1}Z\right) = s_1$$

Therefore the CPE introduction allows user mobile 1 to recover its symbols, immune from the auto-interference and from the interference from other intracell user symbol streams, through simple processing of the received signal, which implies:

a) Discard the *d* samples relative to the cyclic prefix duration in every intersymbol time interval of *n+d* received samples;

b) Take the FFT of **Y** vector to obtain the **Z** vector;

c) Divide frequency-wise the **Z** vector by **S** to equalize the spectral response of the channel so to recover the orthogonality between codes relative to different users; at this point we can follow two different alternatives, indicated with d and e, to obtain the symbol estimate, immune from ISI and MUI.

d) Take the internal product between vectors (= correlation, see (10)) between the spectral transform of the user code 1 and the vector obtained in (c).

e) Inverse transform (from frequency to time) the equalised spectral vector obtained at point c) eq.(9) and then take the internal product between the time code of the user 1 and the inverse transform vector, eq.(11)

The less demanding processing between d) and e) can be implemented.

The eq. (9) indicates that every element of the received spectral vector, (sum of spectral vectors of all users, weighted by the symbols) relative to a certain frequency, is divided by the channel spectral response relative to the same frequency to equalise it from the channel distortion.

A comparison with today MUD processors indicates that CPE is significantly less demanding of processing load Figure 2 summarises the processing in the receiver.

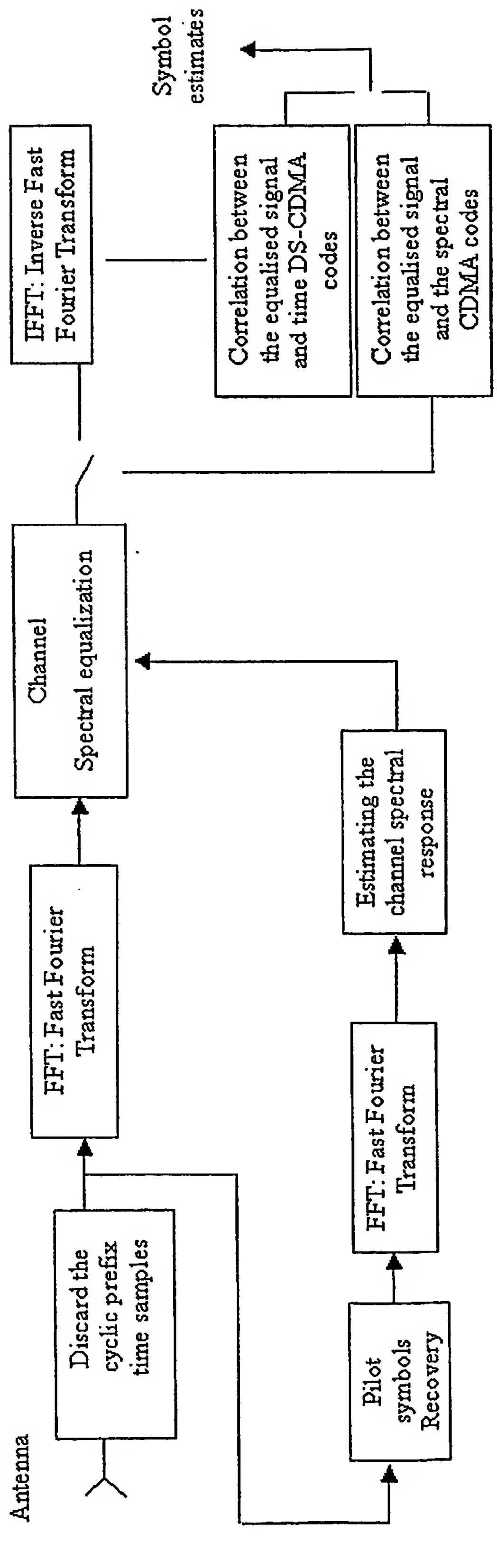

Figura 2: Processing at the mobile to estimate symbols.

1.3.2 Processor for MISO configuration: CPE-2

Background

The aim of orthogonal, Space-Time Block Codes ([4]; [1] fig. 6.15 pag. 97) is to stabilize reception with respect to the statistical fluctuations affecting the radio channel. In this way a gain in signal-noise ratio is achieved by spatial diversity and thus, for the same obtainable performances, we can alternatively transmit less power or, with the same emitted power, extend the cell dimension, which is the maximum distance between a user and the radio base.

Orthogonal Space-Time Block Codes were originally devised, in MISO configuration, for a number of transmitting antennas equal to 2,4,8 [7], for:

- non dispersive time channels (= flat band)
- no DS-CDMA coding for the symbol

We indicate below the space-time processing for the configuration (2,1). We indicate with s(q), s(q+1) symbols to be transmitted in subsequent intersymbol intervals *q*, and *q+1*: - at instant *q* we transmit s(q) from the antenna 1 and s(q+1) from the antenna 2; - at instant *q+1* we transmit the *conjugated* symbols $-s^*(q+1)$ from the antenna 1 and $s^*(q)$ from the antenna 2.

Innovation

The synchronous transmissions from the two transmitting antennas at the radio-base is received through the unique mobile antenna. We indicate by $\mathbf{H}_1$ and $\mathbf{H}_2$ two channel matrices, which are analogous to the one considered in eq. (2) and (3); they are relative to the paths: radio-base antenna 1 – receiving antenna of the user 1, antenna 2 – receiving antenna of the user 1. We generalise both CPE processing for single antenna at the radio base and the Space-Time Block Coding algorithm described in [4] to obtain:

$$\begin{aligned} \mathbf{Y}(q) &= \mathbf{H}_1\left(\mathbf{C}_1 s_1(q) + \mathbf{C}_2 s_2(q)\right) + \mathbf{H}_2\left(\mathbf{C}_1 s_1(q+1) + \mathbf{C}_2 s_2(q+1)\right) \\ \mathbf{Y}(q+1) &= -\mathbf{H}_1\left(\mathbf{G}_1 s_1^*(q+1) + \mathbf{G}_2 s_2^*(q+1)\right) + \mathbf{H}_2\left(\mathbf{G}_1 s_1^*(q) + \mathbf{G}_2 s_2^*(q)\right) \end{aligned} \tag{12}$$

where:

- $s_i(q)$, $s_i(q+1)$: symbols for the *i*-th user, transmitted in the symbol time intervals *q* and (*q+1*); the duration of the intersymbol interval is *n+d* chips;
- $\mathbf{C}_i$ ($\mathbf{G}_i$): DS-CDMA time code of the *i*-th user, to be used in the time interval *q* (*q+1*);

- $\mathbf{Y}$(q) and $\mathbf{Y}$(q+1): time vector received by the user 1, after discarding *the cyclic prefix,* relative to the intervals q and $q+1$ respectively.

In (12), both types of coding, DS-CDMA and orthogonal Space-Time Block Coding, are present.

As a cyclic prefix is adopted, the channel matrix $\mathbf{H}_l$ has a spectral factorisation in eigenvectors and eigenvalues indicated as follows:

(13) $$\mathbf{H}_l = \mathbf{F}\mathbf{S}_l\mathbf{F}^+$$

Similarly as in SISO we define the spectral transforms:

(14) $$\begin{gathered} \mathbf{Z}(\mathrm{q}) = \mathbf{F}^+\mathbf{Y}(\mathrm{q}) \\ \mathbf{Z}(\mathrm{q}+1) = \mathbf{F}^+\mathbf{Y}(\mathrm{q}+1) \end{gathered}$$

(15) $$\begin{gathered} \mathbf{T}_i = \mathbf{F}^+\mathbf{C}_i \\ \mathbf{P}_i = \mathbf{F}^+\mathbf{G}_i \end{gathered}$$

By pre-multiplying by $\mathbf{F}^+$ both members of (12) we have

(16) $$\begin{gathered} \mathbf{Z}(\mathrm{q}) = \mathbf{S}_1\left(\mathbf{T}_1\mathrm{s}_1(\mathrm{q}) + \mathbf{T}_2\mathrm{s}_2(\mathrm{q})\right) + \mathbf{S}_2\left(\mathbf{T}_1\mathrm{s}_1(\mathrm{q}+1) + \mathbf{T}_2\mathrm{s}_2(\mathrm{q}+1)\right) \\ \mathbf{Z}(\mathrm{q}+1) = -\mathbf{S}_1\left(\mathbf{P}_1\mathrm{s}_1^*(\mathrm{q}+1) + \mathbf{P}_2\mathrm{s}_2^*(\mathrm{q}+1)\right) + \mathbf{S}_2\left(\mathbf{P}_1\mathrm{s}_1^*(\mathrm{q}) + \mathbf{P}_2\mathrm{s}_2^*(\mathrm{q})\right) \end{gathered}$$

By taking the conjugate of the last equation we obtain

(17) $$\mathbf{Z}^*(\mathrm{q}+1) = -\mathbf{S}_1^*\left(\mathbf{P}_1^*\mathrm{s}_1(\mathrm{q}+1) + \mathbf{P}_2^*\mathrm{s}_2(\mathrm{q}+1)\right) + \mathbf{S}_2^*\left(\mathbf{P}_1^*\mathrm{s}_1(\mathrm{q}) + \mathbf{P}_2^*\mathrm{s}_2(\mathrm{q})\right)$$

We introduce the following constraint between the spectral transforms of DS-CDMA codes:

(18) $$\mathbf{P}_i^* = \mathbf{T}_i$$

Therefore the spectral codes to be used in the q+1 interval (when symbol conjugates are transmitted) are the conjugate values of those used in the q interval.

By writing the first of the (16) and (17) in matrix notation we have

(19) $$\begin{pmatrix} \mathbf{Z}(\mathrm{q}) \\ \mathbf{Z}^*(\mathrm{q}+1) \end{pmatrix} = \begin{pmatrix} \mathbf{S}_1 & \mathbf{S}_2 \\ \mathbf{S}_2^* & -\mathbf{S}_1^* \end{pmatrix} \cdot \begin{pmatrix} \mathbf{T}_1\mathrm{s}_1(\mathrm{q}) + \mathbf{T}_2\mathrm{s}_2(\mathrm{q}) \\ \mathbf{T}_1\mathrm{s}_1(\mathrm{q}+1) + \mathbf{T}_2\mathrm{s}_2(\mathrm{q}+1) \end{pmatrix}.$$

We define the $\mathbf{S}$ matrix formed by 4 diagonal submatrices, which contain spectral responses relative to the channels between the transmitting antennas 1 and 2 and receiver 1:

(20) $$\mathbf{S} = \begin{pmatrix} \mathbf{S}_1 & \mathbf{S}_2 \\ \mathbf{S}_2^* & -\mathbf{S}_1^* \end{pmatrix}.$$

By pre-multiplying the matrix eq. (19) by the matrix $\mathbf{S}^+$ we obtain

$$(21)\qquad \begin{pmatrix} \mathbf{S}_1^* & \mathbf{S}_2 \\ \mathbf{S}_2^* & -\mathbf{S}_1 \end{pmatrix}\begin{pmatrix} \mathbf{Z}(q) \\ \mathbf{Z}^*(q+1) \end{pmatrix} = \begin{pmatrix} \mathbf{D} & 0 \\ 0 & \mathbf{D} \end{pmatrix}\cdot\begin{pmatrix} \mathbf{T}_1\mathbf{s}_1(q)+\mathbf{T}_2\mathbf{s}_2(q) \\ \mathbf{T}_1\mathbf{s}_1(q+1)+\mathbf{T}_2\mathbf{s}_2(q+1) \end{pmatrix}$$

The matrix **D** is diagonal and has elements:

$$(22)\qquad D_i = |S_{1i}|^2 + |S_{2i}|^2$$

where S_{1i} (S_{2i}) is the i-th element of the spectral response relative to the channel linking the transmitting antenna 1 (2) and the receiver of the user 1.

Thus the first matrix at second member of (21) is also diagonal. By pre-multiplying both members of (21) by the inverse of that matrix we obtain:

$$(23)\begin{pmatrix} \mathbf{D}^{-1} & 0 \\ 0 & \mathbf{D}^{-1} \end{pmatrix}\begin{pmatrix} \mathbf{S}_1^* & \mathbf{S}_2 \\ \mathbf{S}_2^* & -\mathbf{S}_1 \end{pmatrix}\begin{pmatrix} \mathbf{Z}(q) \\ \mathbf{Z}^*(q+1) \end{pmatrix} = \begin{pmatrix} \mathbf{D}^{-1}\mathbf{S}_1^* & \mathbf{D}^{-1}\mathbf{S}_2 \\ \mathbf{D}^{-1}\mathbf{S}_2^* & -\mathbf{D}^{-1}\mathbf{S}_1 \end{pmatrix}\begin{pmatrix} \mathbf{Z}(q) \\ \mathbf{Z}^*(q+1) \end{pmatrix} = \begin{pmatrix} \mathbf{T}_1 s_1(q)+\mathbf{T}_2 s_2(q) \\ \mathbf{T}_1 s_1(q+1)+\mathbf{T}_2 s_2(q+1) \end{pmatrix}$$

Matrix $\mathbf{D}^{-1}$ is diagonal and has elements:

$$(24)\qquad (D^{-1})_i = \frac{1}{|S_{1i}|^2 + |S_{2i}|^2}$$

From the previous matrix equation we write the whole system of equations:

$$(25)\qquad \begin{aligned} \mathbf{T}_1\mathbf{s}_1(q)+\mathbf{T}_2\mathbf{s}_2(q) &= [\mathbf{D}^{-1}\mathbf{S}_1^*]\mathbf{Z}(q) + [\mathbf{D}^{-1}\mathbf{S}_2]\mathbf{Z}^*(q+1) \\ \mathbf{T}_1\mathbf{s}_1(q+1)+\mathbf{T}_2\mathbf{s}_2(q+1) &= [\mathbf{D}^{-1}\mathbf{S}_2^*]\mathbf{Z}(q) - [\mathbf{D}^{-1}\mathbf{S}_1]\mathbf{Z}^*(q+1) \end{aligned}$$

Matrix $\mathbf{D}^{-1}\mathbf{S}_2$ is diagonal and its i-th element is:

$$(26)\qquad (\mathbf{D}^{-1}\mathbf{S}_2)_i = \frac{S_{2i}}{|S_{1i}|^2 + |S_{2i}|^2}.$$

Similarly we define the elements of the other diagonal matrices in the square brackets of eq.s (25).

The processor of the mobile user 1 pre-multiplies the first equation of (25) by $\mathbf{T}_1^+$ to estimate the symbol $s_1(q)$ transmitted to it. We obtain, taking into account (6)

$$(27)\qquad \mathbf{s}_1(q) = \mathbf{T}_1^+[\mathbf{D}^{-1}\mathbf{S}_1^*]\mathbf{Z}(q) + \mathbf{T}_1^+[\mathbf{D}^{-1}\mathbf{S}_2]\mathbf{Z}^*(q+1).$$

By pre-multiplying by $\mathbf{T}_1^+$ the second equation of (25), the user 1 obtains the estimate of symbol $s_1(q+1)$

$$(28)\qquad \mathbf{s}_1(q+1) = \mathbf{T}_1^+[\mathbf{D}^{-1}\mathbf{S}_2^*]\mathbf{Z}(q) - \mathbf{T}_1^+[\mathbf{D}^{-1}\mathbf{S}_1]\mathbf{Z}^*(q+1).$$

The processing described by (27) and (28) extends the spectral processing for DS-CDMA codes considered in (10) to the case of 2 transmitting antennas, 1 receiving antenna and orthogonal space-time block coding; if in (27) we consider only one antenna, and this is equivalent to put S_2=0, the equation reduces to (10). Furthermore,

the processing extends the space-time coding indicated in [4] to the case of DS-CDMA coded signals. In the absence of DS-CDMA codes the scheme indicated above reduces to the known scheme for space-time orthogonal codes with the antenna configuration (1,2). Alternatively to eq. (27) and (28) we can write:

$$\text{(29)} \qquad \mathbf{s}_1(q) = \mathbf{C}_1^+ \left\{ \mathbf{F}\left(\left[\mathbf{D}^{-1}\mathbf{S}_1^* \right] \mathbf{Z}(q) + \mathbf{T}_1^+ \left[\mathbf{D}^{-1}\mathbf{S}_2 \right] \mathbf{Z}^*(q+1) \right) \right\}$$

$$\text{(30)} \qquad \mathbf{s}_1(q+1) = \mathbf{C}_1^+ \left\{ \mathbf{F}\left[\left(\mathbf{D}^{-1}\mathbf{S}_2^* \right) \mathbf{Z}(q) - \mathbf{T}_1^+ \left(\mathbf{D}^{-1}\mathbf{S}_1 \right) \mathbf{Z}^*(q+1) \right] \right\}$$

The equations (27), (28) and (29), (30) indicate that to estimate symbols, correlation between the received (and spectrally equalized) signal and the user code can be implemented either in the frequency domain, with spectral code $\mathbf{T}_1$, or in the time domain (after IFFT) with the time code $\mathbf{C}_1$. Figure 2 indicates the block schemes for the processing at the mobile receiver. When spectral correlation is implemented, the constraint in (18) between spectral codes to be used in intervals q and $q+1$ must be implemented. If a correlation between time codes is implemented, time codes relative to intervals q and $q+1$ must comply with constraint in eq. (31), which is a consequence of the spectral domain constraint indicated in eq. (18).

$$\text{(31)} \qquad \mathbf{G}_i = \mathbf{F}^2\mathbf{C}^*_i$$

$\mathbf{F}^2$ is a real and symmetric *permutation matrix*, having elements equal to

$$\text{(32)} \qquad \left(\mathbf{F}^2\right)_{r,s} = \delta(r+s) \qquad r,s:0,1,\ldots,n-1$$

where

$$\text{(33)} \qquad \delta(r+s) = \begin{cases} 1, \text{if } r+s=0 \text{ or } n \\ 0 \text{ elsewhere} \end{cases}$$

The constraint holds for the transmitter, which transmits time codes.

Matrix $\mathbf{F}^2$ time reverses the positions of the code $\mathbf{C}^*_i$ elements to obtain the time code $\mathbf{G}_i$: the first code element $\mathbf{C}^*_i$ keeps its position, while the positions of the following *n-1* elements are time reverted. Matrix $\mathbf{G}_i$ is equivalent to a time-reversal symbol pre-coding.

1.3.3 MISO CPE processor with potential compatibility among UMTS standard and CPE mobile terminals: CPE-3

To motivate the CPE processor described in this section, we consider a scenario in the cell in which a ME can belong to one of the two following populations, indicated with the subscripts a) and b), characterised by their processing:

- (ME_a): standard processing, for example with Rake. Rake processing in ME_as [1, fig. 37] performs correlation between the user code and delayed replicas of the incoming signal, before coherent recombination of the replicas.
- (ME_b): processing geared around CPE: it is able to work either as a standard ME_a terminal without CPE processing, or with CPE processing. This is required when ME_b roams to standard cells not using CPE.

The basestation sends simultaneously downlink an aggregate signal in which the symbol streams to ME_a do not use CPE, and streams to ME_b use CPE.

Since CPE is a newcomer to the 3G stage, the rationale of the CPE processor indicated below is that the DL streams addressed to ME_bs must not introduce additional MUI to ME_as due to the very presence of CPE transmission. We focus only on a minimal requirement for the less critical scenario, which is the absence of time channel dispersion: MUI from ME_b to ME_a should be absent.

We define the DS-CDMA code matrix **C** as:

$$(34)\qquad \mathbf{C}=\left(\overbrace{\mathbf{C}_{a,1}\ \cdots\ \mathbf{C}_{a,M_a}}^{\mathbf{C}_a}\Big|\overbrace{\mathbf{C}_{b,1}\ \cdots\ \mathbf{C}_{b,M_b}}^{\mathbf{C}_b}\right)$$

where

$\mathbf{C}_{a,k}$ is a code column vector relative to the k^{th} ME_a;

$\mathbf{C}_{b,j}$ is a code column vector relative to the j^{th} ME_b;

$\mathbf{C}_a$ is a code submatrix with M_a columns relative to all the ME_a;

$\mathbf{C}_b$ is a code submatrix with M_b columns relative to all the ME_b.

We define also the symbol vector **s** as

$$(35)\qquad \mathbf{s}=\left(\overbrace{s_{a,1}\ \cdots\ s_{a,M_a}}^{\mathbf{s}_a^T}\Big|\overbrace{s_{b,1}\ \cdots\ s_{b,M_b}}^{\mathbf{s}_b^T}\right)^T$$

where

$s_{a,k}$ is the symbol transmitted to to the k^{th} ME_a;

$s_{b,j}$ is the symbol transmitted to to the j^{th} ME_b;

s_a is a symbol column vector with M_a rows comprising the symbols to all the ME_a;

s_b is a symbol column vector with M_b rows comprising the symbols to all the ME_b.

By considering first the same CPE processor as in eq. (12) we can write

$$(36)\qquad \begin{aligned} &\mathbf{Y}(q)=\mathbf{H}_1\left(\mathbf{C}_a\mathbf{s}_a(q)+\mathbf{C}_b\mathbf{s}_b(q)\right)+\mathbf{H}_2\left(\mathbf{C}_a\mathbf{s}_a(q+1)+\mathbf{C}_b\mathbf{s}_b(q+1)\right)\\ &\mathbf{Y}(q+1)=-\mathbf{H}_1\left(\mathbf{C}_a\mathbf{s}_a^*(q+1)+\mathbf{G}_b\mathbf{s}_b^*(q+1)\right)+\mathbf{H}_2\left(\mathbf{C}_a\mathbf{s}_a^*(q)+\mathbf{G}_b\mathbf{s}_b^*(q)\right)\end{aligned}$$

where the time reversed $\mathbf{G}_b$ matrix, in analogy with (31), is defined as

$$(37)\qquad \mathbf{G}_b=\mathbf{F}^2\mathbf{C}^*_b$$

In the absence of time spread (i.e. $\mathbf{H}_1=\mathbf{H}_2=\mathbf{I}_n$) $\mathrm{ME}_{a,1}$ premultiplies the first and second equation of (36) by $\mathbf{C}^+_{a,1}$ to recover its symbol stream

$$(38)\qquad \begin{aligned} &\mathbf{C}^+_{a,1}\mathbf{Y}(q)=s_{a,1}(q)+s_{a,1}(q+1)\\ &\mathbf{C}^T_{a,1}\mathbf{Y}^*(q+1)=-s_{a,1}(q+1)+s_a(q)-\left(\mathbf{C}^T_{a,1}\mathbf{F}^2\mathbf{C}_b s_b(q+1)\right)+\left(\mathbf{C}^T_{a,1}\mathbf{F}^2\mathbf{C}_b s_b(q)\right)\end{aligned}$$

The $\mathrm{ME}_{a,1}$ code $\mathbf{C}_{a,1}$ is not orthogonal to the time reversed code matrix $\mathbf{F}^2\mathbf{C}^*_b$, so MUI arises from symbol stream addressed to ME_bs.

To eliminate this MUI and taking into account (13) we modify (36) as follows:

$$(39)\qquad \begin{aligned} &\mathbf{Y}(q)=\mathbf{F}\mathbf{S}_1\mathbf{F}^+\left(\mathbf{C}_a\mathbf{s}_a(q)+\mathbf{C}_b\mathbf{s}_b(q)\right)+\mathbf{F}\mathbf{S}_2\mathbf{F}^+\left(\mathbf{C}_a\mathbf{s}_a(q+1)+\mathbf{C}_b\mathbf{s}_b(q+1)\right)\\ &\mathbf{Y}(q+1)=-\mathbf{F}\mathbf{S}_1\mathbf{F}^+\left(\mathbf{C}_a\mathbf{s}_a^*(q+1)+\mathbf{C}_b\mathbf{K}_b\mathbf{s}_b^*(q+1)\right)+\mathbf{F}\mathbf{S}_2\mathbf{F}^+\left(\mathbf{C}_a\mathbf{s}_a^*(q)+\mathbf{C}_b\mathbf{K}_b\mathbf{s}_b^*(q)\right)\end{aligned}$$

$\mathbf{K}_b$ is a matrix, whose structure will be determined in the following.

In the absence of time spread (i.e. $\mathbf{H}_1=\mathbf{H}_2=\mathbf{I}_n$), $\mathrm{ME}_{a,1}$ premultiplies the first and second equation by $\mathbf{C}^+_{a,1}$ to recover its symbol stream, to obtain

$$(40)\qquad \begin{aligned} &\mathbf{C}^+_{a,1}\mathbf{Y}(q)=s_{a,1}(q)+s_{a,1}(q+1)\\ &\mathbf{C}^T_{a,1}\mathbf{Y}^*(q+1)=-s_{a,1}(q+1)+s_a(q)-\left(\mathbf{C}^T_{a,1}\mathbf{C}_b\mathbf{K}_b s_b(q+1)\right)+\left(\mathbf{C}^T_{a,1}\mathbf{C}_b\mathbf{K}_b s_b(q)\right)\end{aligned}$$

In this case, independently from the structure of $\mathbf{K}_b$, MUI from ME_b terminals vanishes due to code orthogonality.

Consider now the CPE processing in $\mathrm{ME}_{b,1}$ for recovering its symbols. Premultiplying (39) by $\mathbf{F}^+$, we have

$$(41)\qquad \begin{aligned} &\mathbf{F}^+\mathbf{Y}(q)=\mathbf{S}_1\mathbf{F}^+\left(\mathbf{C}_a\mathbf{s}_a(q)+\mathbf{C}_b\mathbf{s}_b(q)\right)+\mathbf{S}_2\mathbf{F}^+\left(\mathbf{C}_a\mathbf{s}_a(q+1)+\mathbf{C}_b\mathbf{s}_b(q+1)\right)\\ &\mathbf{F}^T\mathbf{Y}^*(q+1)=-\mathbf{S}_1^*\mathbf{F}^T\left(\mathbf{C}_a^*\mathbf{s}_a(q+1)+\mathbf{C}_b^*\mathbf{K}_b^*\mathbf{s}_b(q+1)\right)+\mathbf{S}_2^*\mathbf{F}^T\left(\mathbf{C}_a^*\mathbf{s}_a(q)+\mathbf{C}_b^*\mathbf{K}_b^*\mathbf{s}_b(q)\right)\end{aligned}$$

we introduce the following constraint:

$$(42)\qquad \mathbf{F}^T\mathbf{C}^*_b\mathbf{K}^*_b=\mathbf{F}^+\mathbf{C}_b$$

from which we obtain, taking into account that $\mathbf{F}$ is symmetric,

$$(43)\qquad \mathbf{K}_b=\mathbf{C}^+_b\mathbf{F}^2\mathbf{C}^*_b$$

By doing so, we obtain

$$(44)\qquad \begin{aligned} &\mathbf{F}^+\mathbf{Y}(q)=\mathbf{S}_1\mathbf{F}^+\left(\mathbf{C}_a\mathbf{s}_a(q)+\mathbf{C}_b\mathbf{s}_b(q)\right)+\mathbf{S}_2\mathbf{F}^+\left(\mathbf{C}_a\mathbf{s}_a(q+1)+\mathbf{C}_b\mathbf{s}_b(q+1)\right)\\ &\mathbf{F}^T\mathbf{Y}^*(q+1)=-\mathbf{S}_1^*\left(\mathbf{F}^T\mathbf{C}_a^*\mathbf{s}_a(q+1)+\mathbf{F}^+\mathbf{C}_b\mathbf{s}_b(q+1)\right)+\mathbf{S}_2^*\left(\mathbf{F}^T\mathbf{C}_a^*\mathbf{s}_a(q)+\mathbf{F}^+\mathbf{C}_b\mathbf{s}_b(q)\right)\end{aligned}$$

Following the notation (19) we have

(45) $$\begin{pmatrix} \mathbf{Z}(q) \\ \mathbf{Z}^*(q+1) \end{pmatrix} = \begin{pmatrix} \mathbf{S}_1 & \mathbf{S}_2 \\ \mathbf{S}_2^* & -\mathbf{S}_1^* \end{pmatrix} \begin{pmatrix} \mathbf{F}^+\mathbf{C}_b\mathbf{s}_b(q) \\ \mathbf{F}^+\mathbf{C}_b\mathbf{s}_b(q+1) \end{pmatrix} \text{ plus terms due to } \mathrm{ME_as}$$

If ME_a terminals were absent, a processing as the one indicated after (19) would be valid to estimate the $ME_{a,1}$ symbols.

This processing can be actually applied ignoring the MUI interference on ME_bs on ME_as. This MUI can be coped by other means, not indicated here. We note that the terms in (45) due to ME_as are the same irrespectively of the choice of the symbol pre-coders (37) or (43). Then the latter is preferable as it minimizes MUI from ME_{bs} to ME_as .

The term $\mathbf{C}_b\mathbf{K}_b\mathbf{s}^*_b(q)$ which appears in (39) indicates that instead of the symbol vector $\mathbf{s}^*_b(q)$, the pre-coded symbol vector $\mathbf{K}_b\mathbf{s}^*_b(q)$ is transmitted. The pre-coding matrix $\mathbf{K}_b$ is not time-reversal.

1.3.4 Generalisations of the processors

- *To 4 and 8 transmitting antennas.*

The processing for 2 antennas at the radio base transmitter and 1 antenna at the receiver can be straightforwardly generalised to 4 and 8 transmit antennas at the radio base, by concatenating the generalised STBC codes indicated in [7] with DS-CDMA codes, as it has been done in equations (12) to (16) or (39). In the inter-symbol interval in which in generalised STBC codes the conjugate symbols are transmitted, the symbol pre-coding, respectively indicated in (37) for CPE-2 and in (43) for CPE-3 has to be applied.

- *To take into account of noise correlation matrix*

Processors considered so far, which do not take into account the noise correlation matrix, are indicated as Zero Forcing (ZF). When noise correlation matrix is taken into account, other CPE processors, such as Minimum Mean Square Estimators (MMSE), whose structure is immediate to derive [6], might be considered instead of ZF.

1.3.5 Cyclic prefix and multirate

So far it has been assumed that, after discarding d samples relative to the cyclic prefix, the number n of samples in the FFT time window, is equal to the number of chips of the DS-CDMA code. A milder constraint can be assumed, which is the FFT time window is taken equal to an integer number (r) of DS-CDMA symbols, where r is a parameter that can be freely selected to optimise performance. Now the FFT time window contains rn chips. After equalization and Inverse Fast Fourier Transform (IFFT: from frequency to

time) the *rn* chip equalized vector is segmented in not overlapping sections with durations equal to the code length, and correlation processing with time codes can be implemented. In the multirate case [1] the radio-base transmits an aggregate sum of symbol streams, with codes of different lengths (for example, in case of OVSF codes: Orthogonal Variable Spreading Factor) for different users. For the aggregate sum a window is taken whose length is an integer multiple of the duration of the longest code, which is cyclically extended before transmission. In reception, after discarding cyclic prefix, the received vector is FFT transformed, spectrally equalized, IFFT inverse transformed and segmented as indicated above before time correlation; each user has a correlation window equal to its code length.

1.3.6 Iterative Spectral Processing for SISO: CPE-4

The matrix **H** in eq. (3) is the sum of 2 matrices: $\mathbf{H}_0$, $\mathbf{H}_1$

(46) $$\mathbf{H}_0 = \begin{pmatrix} h_0 & 0 & 0 & 0 & 0 \\ h_1 & h_0 & 0 & 0 & 0 \\ h_2 & h_1 & h_0 & 0 & 0 \\ 0 & h_2 & h_1 & h_0 & 0 \\ 0 & 0 & h_2 & h_1 & h_0 \end{pmatrix}$$

(47) $$\mathbf{H}_1 = \begin{pmatrix} 0 & 0 & 0 & h_2 & h_1 \\ 0 & 0 & 0 & 0 & h_2 \\ 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 \end{pmatrix}$$

By indicating with indexes *m*, *m-1* the time intervals associated to subsequent time symbols, the time vector of the samples received in m^{th} step is $\mathbf{Y}_m$, whereas the two users' symbols in the two intervals are indicated with $s_{1,m}$, $s_{2,m}$, $s_{1,m\text{-}1}$, $s_{2,m\text{-}1}$. In the absence of cyclic prefix we have

(48) $$\begin{aligned} \mathbf{Y}_m &= \mathbf{H}_0(\mathbf{C}_1 s_{1,m} + \mathbf{C}_2 s_{2,m}) + \mathbf{H}_1(\mathbf{C}_1 s_{1,m-1} + \mathbf{C}_2 s_{2,m-1}) \\ &= \mathbf{H}(\mathbf{C}_1 s_{1,m} + \mathbf{C}_2 s_{2,m}) - \mathbf{H}_1(\mathbf{C}_1 s_{1,m} + \mathbf{C}_2 s_{2,m}) + \mathbf{H}_1(\mathbf{C}_1 s_{1,m-1} + \mathbf{C}_2 s_{2,m-1}) \end{aligned}$$

We assume that the symbols in the (*m-1*)th interval are known and can be cancelled from eq. (48);

thus:

(49) $$\mathbf{Y}_m = \mathbf{H}(\mathbf{C}_1 s_{1,m} + \mathbf{C}_2 s_{2,m}) - \mathbf{H}_1(\mathbf{C}_1 s_{1,m} + \mathbf{C}_2 s_{2,m})$$

By applying the spectral transform in the same way as indicated in eq. (8) we obtain

(50) $$\mathbf{Z} = \mathbf{S}(\mathbf{T}_1 s_{1,m} + \mathbf{T}_2 s_{2,m}) - \mathbf{Q}(\mathbf{T}_1 s_{1,m} + \mathbf{T}_2 s_{2,m})$$

where the **Q** matrix is given by

(51) $$\mathbf{Q} = \mathbf{F}^{+}\mathbf{H}_1\mathbf{F}$$

We indicate with $\mathrm{abs}(x_m)$ and $\min_m(x_m)$ the absolute and minimum value of the quantity x_m taken with respect to the index m.

By indicating with N_{it} the maximum number of iterations and with c the maximum acceptable error on the estimate of the symbol, and by indicating with $\hat{s}_{i,m}(k)$ the ith user symbol estimate in the kth step, an iterative estimate of the two symbols $s_{1,m}$, $s_{2,m}$ is carried out as follows:

In step 0 we set $\hat{s}_{i,m}(0) = 0$ and $k=1$;

$\mathbf{Z}_m(k) = \mathbf{Z}_m + \mathbf{Q}(\mathbf{T}_1 \hat{s}_{1,m}(k-1) + \mathbf{T}_2 \hat{s}_{2,m}(k-1))$; $\hat{s}_{i,m}(k) = \mathbf{T}_i^{+}\mathbf{S}^{-1}\mathbf{Z}_m(k)$; $k=k+1$; If $k \le N_{it}$ and $\min_m(\mathrm{abs}(\hat{s}_{i,m}(k) - \hat{s}_{i,m}(k-1))) > c$, then repeat step 2; else the iterative process is over. The iteration process stops either when k attains the value N_{it} or when the minimum indicated above becomes less than the value c. The iterative spectral estimation performs less well than the symbol estimation when using cyclic prefix, it has greater computational load but it does not require the introduction of a cyclic prefix.

1.3.7 References

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1.4 Best mode for carrying out the invention

- Assessing in detail the system advantages achievable by CPE processing with respect to standard 3G techniques.

- Assess the modification to the standards required by CPE techniques; in particular, exploit the greater potential compatibility of CPE-3 with standard terminals to ease the acceptance by the standardization boards.

- Submit the CPE techniques to standardization boards such as 3GPP as optional techniques.

- Promote CPE techniques to big manufacturers.

1.5 Industrial applicability

Straightforward deducible from the disclosure of the invention.

1.6 CLAIMS

1. **CLAIM 1**: Synthesis of a new spectral processing method minimising Inter Symbol self Interference (ISI) and Multi Users Interference (MUI), named CPE-1 and detailed in section 1.3.1), based on the introduction of a cyclic prefix for communications systems using DS-CDMA time coding and for one antenna element at both mobile receiver and base station transmitter. Such processing has two alternative methods of carrying out time or frequency correlations.
2. **CLAIM 2**: Synthesis of a new spectral processing method minimising Inter Symbol self Interference (ISI) and Multi Users Interference (MUI), named CPE-2 and detailed in section 1.3.2, based on the introduction of a time-reversal symbol pre-coding and a cyclic prefix for communications systems using DS-CDMA time coding and for two antenna elements at base station transmitter and one antenna element at mobile receiver.
3. Using the same processing structure as reported in CLAIM 2, but for 4 or 8 antenna elements at the transmitting base station and one antenna element at the mobile receiver.
4. **CLAIM 3**: Synthesis of a new spectral processing method minimising Inter Symbol self Interference (ISI) and Multi Users Interference (MUI), named CPE-3 and detailed in section 1.3.3, based on symbol non-time-reversal pre-coding at the transmitter and on the introduction of a cyclic prefix for communications systems using DS-CDMA time coding and for two antenna elements at base station transmitter and one antenna element at mobile receiver.
5. Using the same processing structure as reported in CLAIM 3, but for 4 or 8 antenna elements at the transmitting base station and one antenna element at the mobile receiver.
6. **CLAIM 4**: Synthesis of a new spectral processing method minimising Inter Symbol self Interference (ISI) and Multi Users Interference (MUI), without cyclic prefix introduction, named CPE-4 and detailed in section 1.3.6, for communications systems using DS-CDMA time coding, to estimate the sequence of symbols arriving at the mobile receiver. Such method is valid for systems employing 1 antenna element both at the base station transmitter and at mobile receiver.